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                A PHASE CORRECTION SYSTEM TO IMPROVE THE QUALITY OF A WIDEBAND --ETC(U)
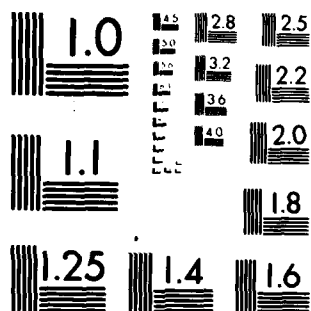   FEB 80   B LOESCH   F19628-80-C-0002
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LINCOLN LABORATORY

A PHASE CORRECTION SYSTEM TO IMPROVE THE QUALITY OF A WIDEBAND CHIRP SIGNAL

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ABSTRACT

A phase correction circuit which significantly improves
the linearity of a wideband (hundreds of MHz) chirp signal has
been constructed and tested. This correction circuit is appli-
cable to a chirp generator in which a digital system is used to
compute the phase of the chirp signal once each clock cycle,
and the chirp signal bandwidth is then restricted to appreciably
less than the clock frequency; analog frequency multiplication
is then used to obtain the required bandwidth. The circuit
described here corrects the high frequency phase errors which
are present after the multiplication process.

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I. INTRODUCTION

One method of generating chirp signals with hundreds of MHz bandwidth is to use a digital system to compute the phase of the chirp signal once each clock cycle and to then use analog frequency multiplication to obtain the required bandwidth. This multiplication is necessary because the digitally derived bandwidth is restricted to appreciably less than the clock frequency. With this method of wideband signal generation, it has been found that upon correlation, the close-in sidelobes are low because the low frequency components of the digitally derived signal are accurate. Unfortunately, however, the digital phase "noise" produces far-out sidelobes, and the frequency multiplication process increases this phase noise so that these sidelobes reach an undesirably high level.

In a memorandum dated 19 July 1979, it was proposed that after frequency multiplication, the undesired phase "noise" might be detected and then connected to a phase modulator so as to cancel this undesired "noise". This report describes the design and the test results which have been obtained on such a circuit.

II. CHIRP SIGNAL GENERATOR AND CORRELATOR

The tests in this report were made using a signal generator and a correlator which were available. This generator was fixed in design and was constructed as follows:

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From two read-only memories digital stored sine and cosine baseband values of the chirp signal were read out at a 50 MHz clock rate. These values were such as to produce contiguous chirp signals, 10 microseconds in length, and with ±15.6 MHz bandwidth. D/A converters and specially designed deglichers were then used to convert the digital signals to analog form. This pair of baseband signals was then used to drive a single-sideband-modulator operating with a carrier frequency of 120 MHz. The output of this modulator was then passed through successive frequency-doubling and filtering stages so that after a times 16 frequency multiplication, the output signal consisted of contiguous 10 microsecond chirp pulses centered at 1.92 GHz and with 500 MHz bandwidth.

The generator was designed to run with a 10 millisecond period - 5.0 milliseconds on (500 contiguous pulses), and 5.0 milliseconds off. All frequencies in the generator were coherently derived from one accurate 10 MHz crystal oscillator.

The test correlator system consisted of two channels with ±50 and ±100 nanoseconds relative delays and an offset of 200 MHz in one channel. Thus after combining channels, the output signal consisted of 10 microsecond CW pulses centered at 200 MHz for zero delay, and with frequency deviation proportional to relative delay. These pulses were then "weighted" with a cosine-on-a-pedestal in an active modulator so that the final output as viewed on a spectrum analyzer would show one main frequency output with low residual sidelobe response.

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III. PHASE CANCELLATION CIRCUIT

There does not appear to be a practical means of measuring the phase noise (deviation from quadratic) directly. Instead, the deviation from linear frequency is measured by the use of a wide-band frequency discriminator constructed as shown in Fig. 1a. However, the major output from such a discriminator will, of course, be a sawtooth waveform proportional to the frequency sweep of the chirp signal, and it is only the deviations from a linear sawtooth which should be detected and cancelled.

In order to cancel the linear term of the frequency discriminator output, a video sawtooth generator of the opposite sense sawtooth was constructed. This circuit is shown in Fig. 1b. It receives a 2.5 volt positive pulse approximately 100 nsec wide at the chirp snap-back time and the level and amplitude of the output sawtooth are adjusted for best cancellation of the discriminator sawtooth output.

In Fig. 2a is shown the frequency discriminator output with the chirp signal input only. In Fig. 2b is shown the sum signal with the linear video compensation circuit output added.

The signal shown in Fig. 2b represents frequency deviation from a linear ramp. This must now be integrated so as to represent phase deviation, and also be amplified to the proper level to drive the phase modulator. A schematic diagram of the six stage amplifier to accomplish these requirements is shown in
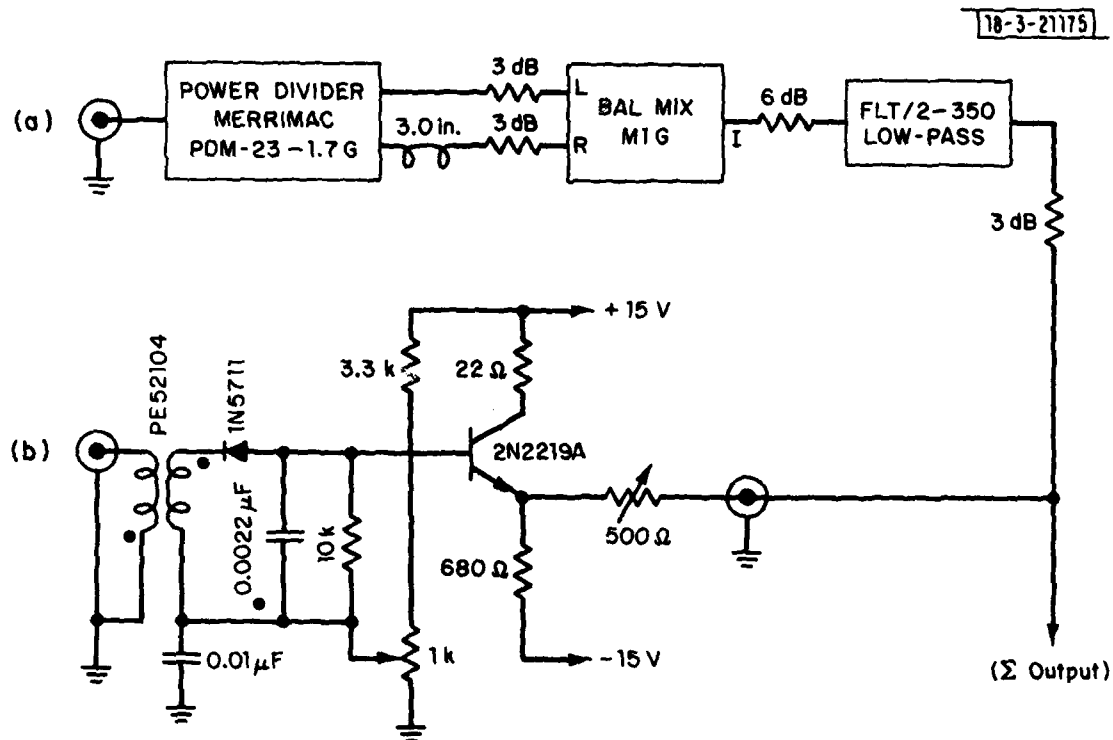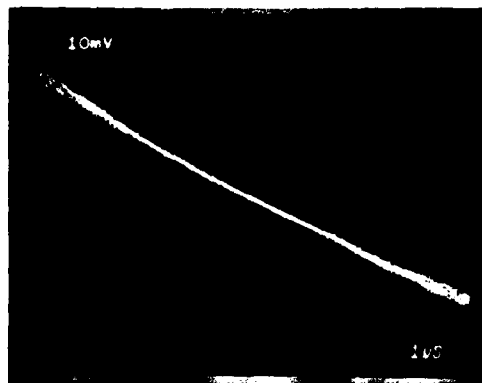
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(a) POWER DIVIDER
MERRIMAC
PDM-23-1.7G

3.0 in.

3 dB

3 dB

L

R

BAL MIX
M1G

I

6 dB

FLT/2-350
LOW-PASS

3 dB

(b) PE52104

1N5711

0.0022 μF

0.01 μF

10k

1k

3.3 k

22 Ω

+15 V

2N2219A

680 Ω

-15 V

500 Ω

(Σ Output)

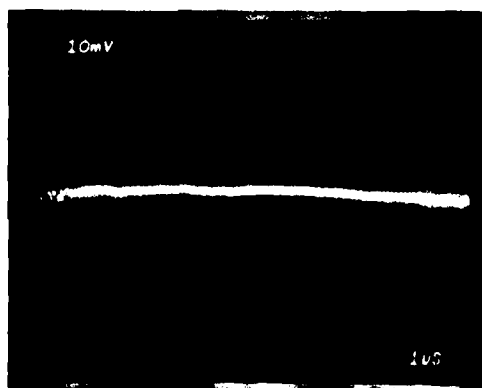Fig. 1.   (a) Wideband frequency discriminator.
(b) Video sawtooth generator.

Fig. 2. (a) Frequency discriminator output with 10 μsec, 500 MHz, chirp input signal. (b) Frequency discriminator plus video sawtooth output.

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Fig. 3. The first two coupling capacitors (.004 μF and .008 μF respectively) are purposely made small so as to restrict the low frequency response; the 6.8 microhenry series choke provides approximate integration for frequencies above 5.0 MHz. There is also a shunt resonant trap tuned to approximately 2.0 MHz to reduce amplifier gain at this frequency. All of these elements have been partly empirically determined for best sidelobe cancellation.

This amplifier output is shown in Fig. 4. It may be noted that for about the first two microseconds of the pulse, the amplifier is operating at or near saturation because the step of the video sawtooth does not accurately enough cancel the step of the frequency discriminator output. The signal shown in this figure is connected to the phase modulator.

The delay through the phase amplifier was measured to be approximately 10 nanoseconds and therefore a compensating coax cable with this delay was used to connect the chirp signal into the phase modulator. This modulator is of standard design as shown in Fig. 5. It was necessary, however, to accurately match the delay through the balanced mixer with a corresponding length of coax line so that the modulation sideband would produce true phase modulation.
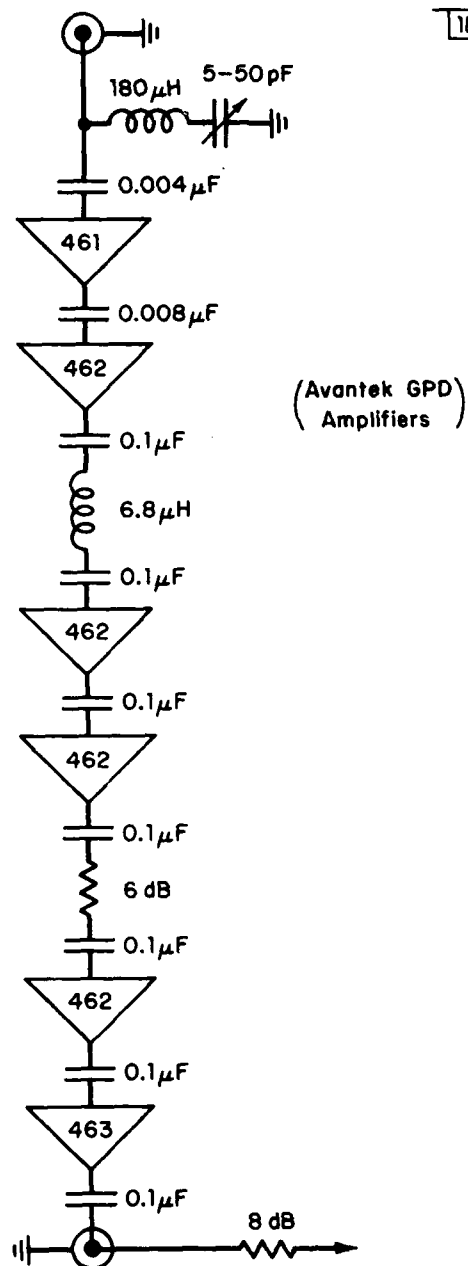
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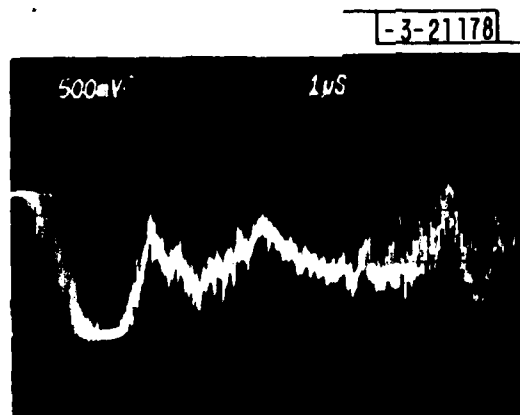Fig. 3.   Schematic diagram of phase correction
amplifier.

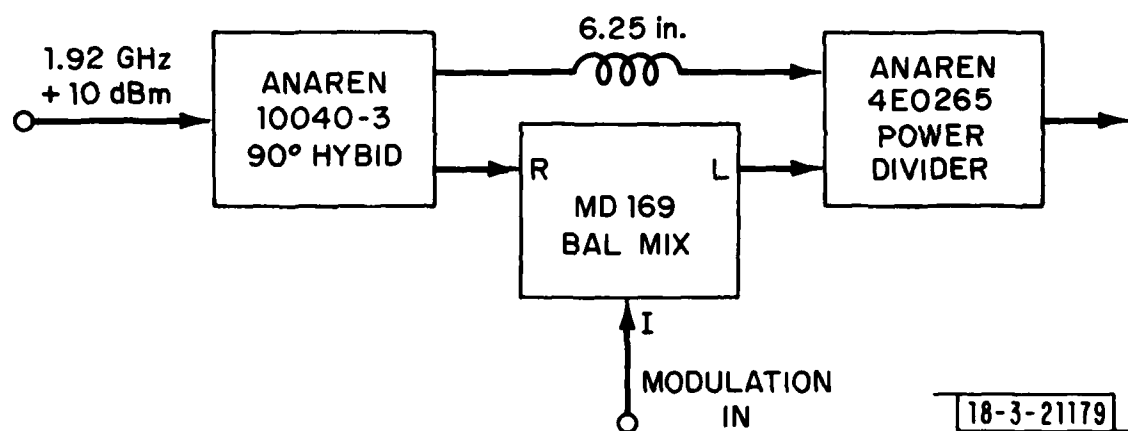Fig. 4. Phase correction amplifier output
with chirp input.

Fig. 5. Schematic diagram of phase modulator.

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IV. SYSTEM TESTS

A time picture of the correlated signal output is shown in Fig. 6. Although it cannot be determined from this picture the carrier frequency in this picture is at about 200 MHz. The envelope of this picture is not discernibly changed with or without phase cancellation applied. This picture was taken with 50 nanoseconds correlation delay and a slight gap can be seen at the minimum values of the envelope. Other than the change in this gap, this picture appears unchanged under any test conditions described in this report. This fact is mentioned because it shows that essentially the complete 10 microsecond pulse always appears as correlated output, and thus the doppler resolution is not changed by phase correction.

In Fig. 7 is shown a picture from a spectrum analyzer of the correlated signal with zero relative delay. This picture shows very low sidelobes and simply shows that the correlation circuit works correctly. This picture does not change with phase correction because non-linear frequency of the chirp signal does not produce any error at zero delay.

In Figs. 8 through 11 are shown spectrum analyzer pictures of the correlated outputs for delays of ±50 and ±100 nanoseconds with and without phase correction. In each figure the top picture was taken without phase correction and the bottom picture was taken with phase correction. In each case the side-lobe level is considerably reduced with phase correction, from
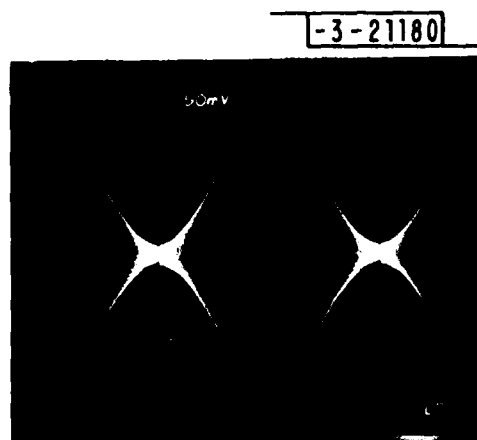
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Fig. 6.   Correlated weighted signal with
50 nsec delay.

Fig. 7.   Spectrum of correlated, weighted signal
with zero delay; 5.0 MHz/cm - horizontal,
200 MHz - center, 10 dB/cm - vertical, 100 KHz -
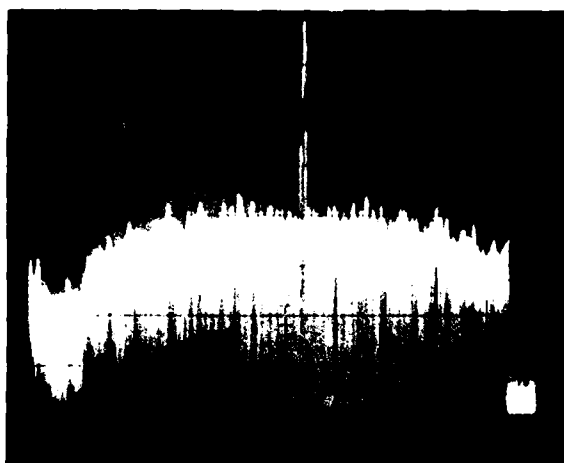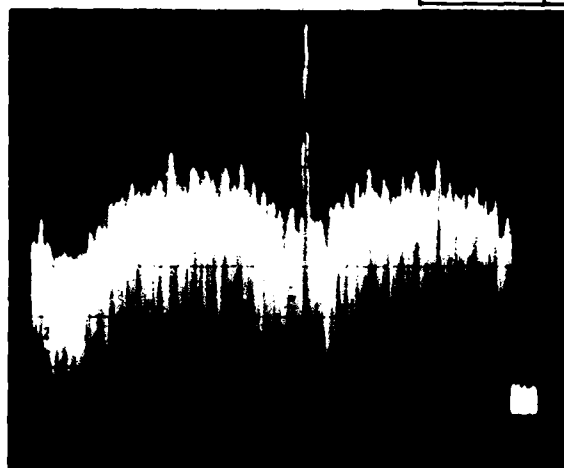resolution.

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Fig. 8. Spectrum of correlated weighted signal with
+50 nsec delay; top - no phase correction, bottom -
with phase correction; 5.0 MHz/cm - horizontal,
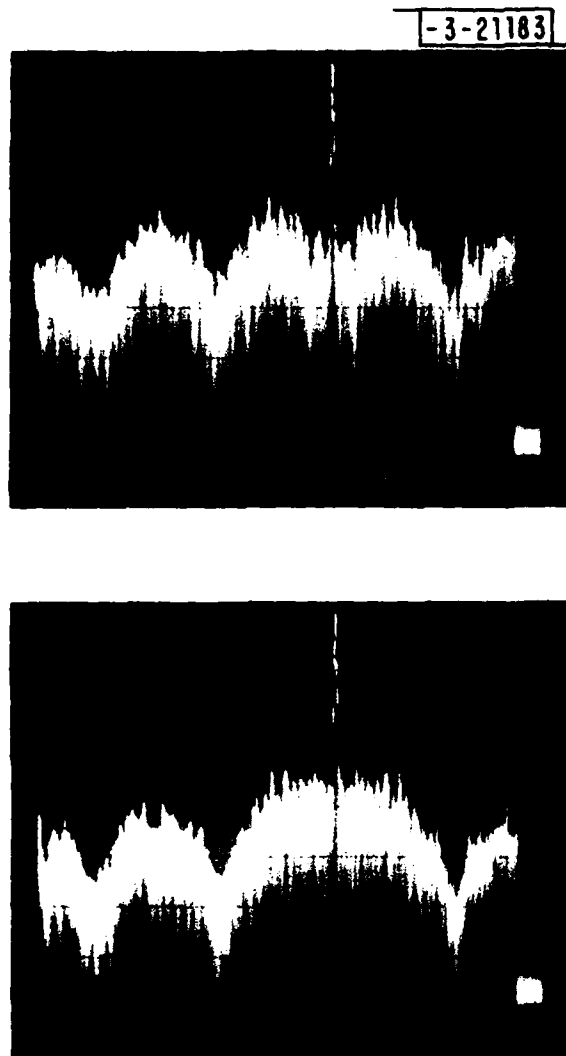200 MHz - center, 10 dB/cm - vertical, 100 KHz -
resolution.

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Fig. 9. Spectrum of correlated weighted signal with
+100 nsec delay; top - no phase correction, bottom -
with phase correction; 5.0 MHz/cm - horizontal,
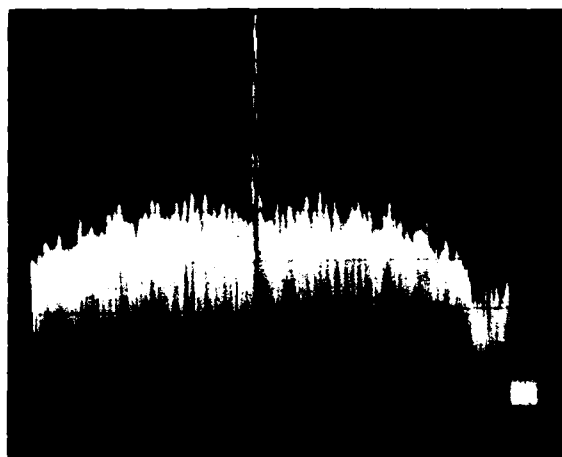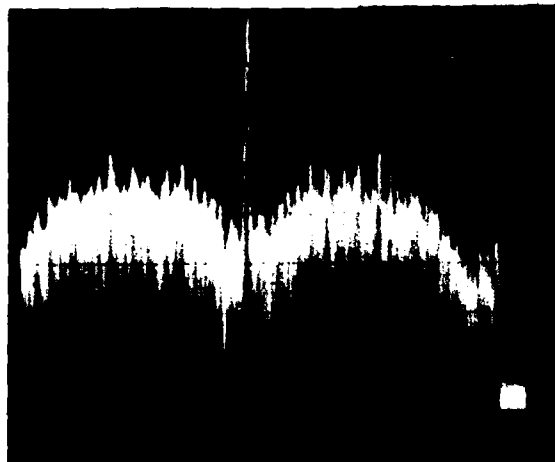200 MHz - center, 10 dB/cm - vertical, 100 KHz -
resolution.

Fig. 10. Spectrum of correlated weighted signal with
-50 nsec delay; top - no phase correction, bottom -
with phase correction; 5.0 MHz/cm - horizontal,
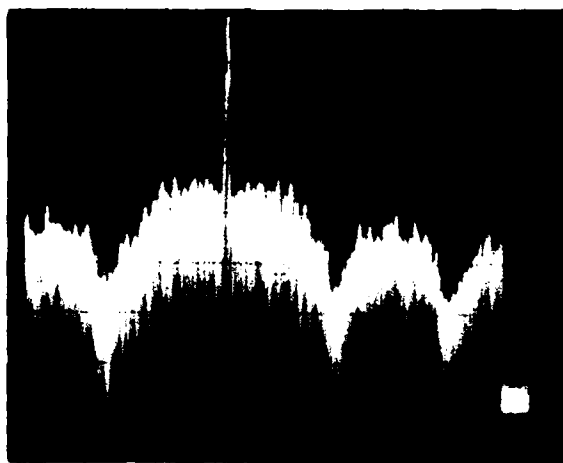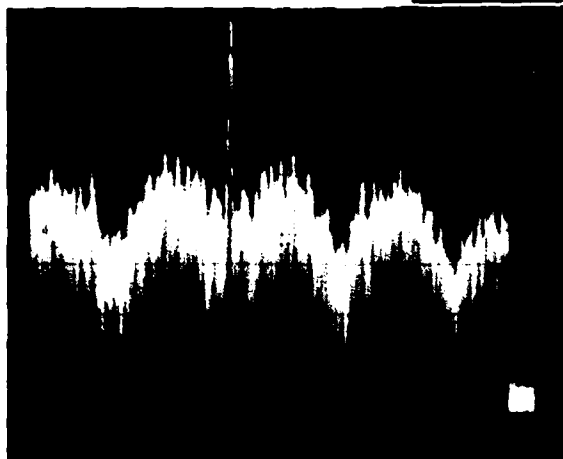200 MHz - center, 10 dB/cm - vertical, 100 KHz -
resolution.

Fig. 11. Spectrum of correlated weighted signal with
-100 nsec delay; top - no phase correction, bottom -
with phase correction; 5.0 MHz/cm - horizontal,
200 MHz - center, 10 dB/cm - vertical, 100 KHz -
resolution.

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a peak level of -28 dB to about -36 dB at ±50 nanoseconds delay and from a peak level of -28 dB to about -34 dB at ±100 nanoseconds.

Figure 12 shows a spectrum analyzer picture of the main lobe output for - 50 nanoseconds delay with a much expanded horizontal scale (50 KHz/cm instead of 5 MHz/cm) and with a linear rather than logarithmic vertical scale; the analyzer resolution was set at 3.0 KHz rather than 100 KHz. This picture is indicative of the doppler resolution of the 10 microsecond chirp signal, but it does not well define this resolution because the pulses are contiguous with a resultant 100 KHz spectral line spacing. (The -3.0 dB spectral width of a Hamming weighted 10 microsecond pulse should be 133 KHz.) This picture, however, is unchanged with or without phase correction.

The range resolution of a waveform depends basically upon bandwidth, and the phase correction system must therefore not reduce the chirp bandwidth. In Fig. 13 is shown the chirp signal spectrum without phase correction (top) and with phase correction (bottom). The spectrum is essentially unchanged being 500 MHz wide in each picture.

V. CONCLUSION

A phase correction system which corrects the phase deviations on a linear frequency modulated chirp signal has been constructed and tested. A significant improvement in the quality
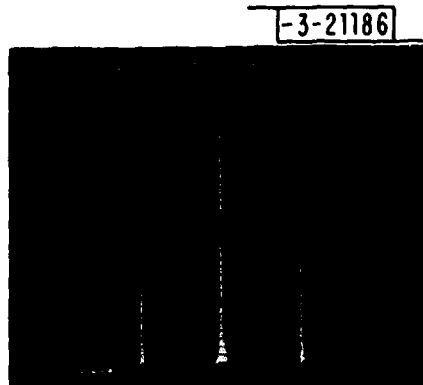
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Fig. 12.   Spectrum of correlated weighted signal with
-50 nsec delay with phase correction; 50 KHz/cm -
horizontal, 198 MHz - center, linear vertical, 3.0 KHz
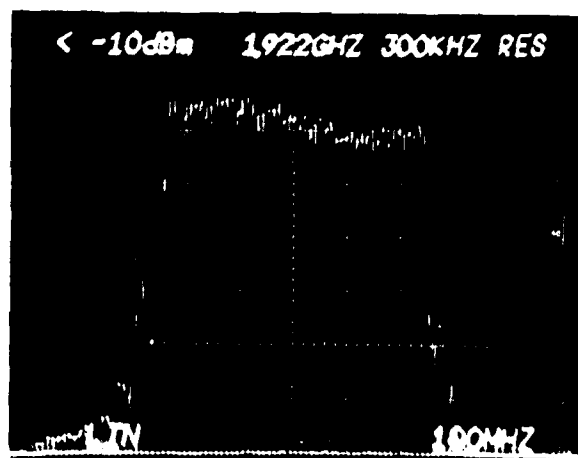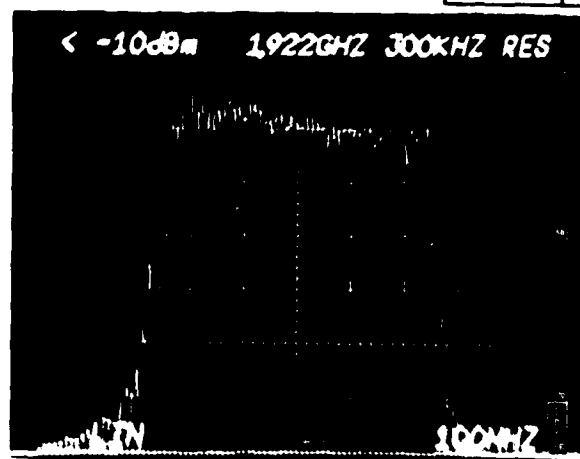- resolution.

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Fig. 13. Spectrum of chirp signal; top - no phase
correction, bottom - with phase correction.

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of a wideband chirp signal has been demonstrated. This correction system may be made better by the use of a more exact video sawtooth generator which may then yield better phase correction over the complete length of the chirp signal.

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A phase correction circuit which significantly improves the linearity of a wideband (hundreds of MHz) chirp signal has been constructed and tested. This correction circuit is applicable to a chirp generator in which a digital system is used to compute the phase of the chirp signal once each clock cycle, and the chirp signal bandwidth is then restricted to appreciably less than the clock frequency; analog frequency multiplication is then used to obtain the required bandwidth. The circuit described here corrects the high frequency phase errors which are present after the multiplication process.

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